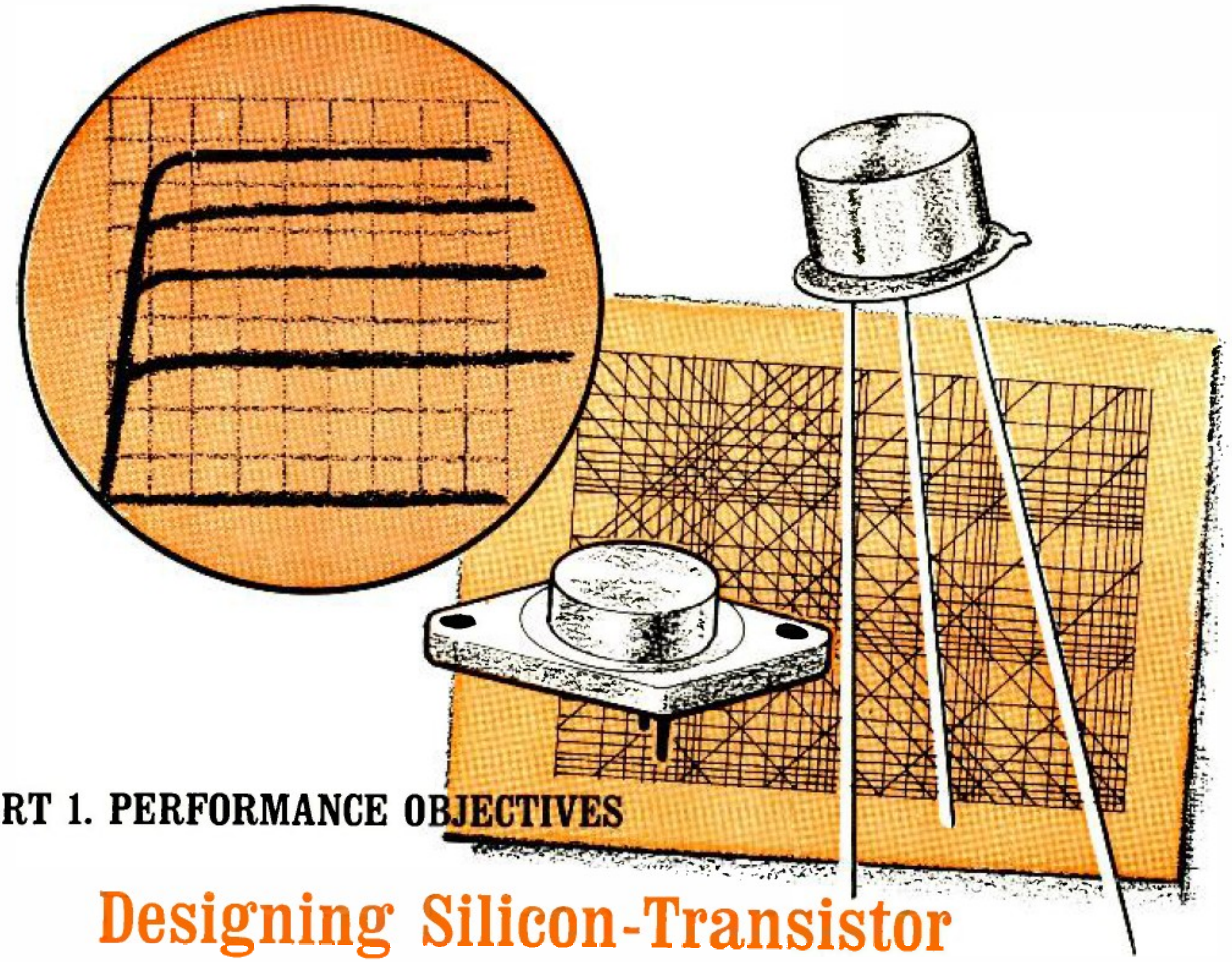


PART 1. PERFORMANCE OBJECTIVES

Designing Silicon-Transistor Hi-Fi Amplifiers

Silicon power transistors now available offer a number of advantages for power output and driver stages. This series of three articles covers the considerations for conservative design and the performance of three practical hi-fi circuits.

By R. D. GOLD and J. C. SONDERMEYER
RCA Electronic Components & Devices

SILICON power transistors that are now available offer many advantages when used in the power-output and driver stages of high-power audio amplifiers, as well as in applications such as servomechanism control systems and high-power ultrasonic generators. This three-part series explains the advantages of using such transistors in high-quality audio systems designed to deliver tens of watts of audio output power and to operate over a wide range of ambient temperatures. The discussion will deal primarily with the power-amplifier and driver stages of such systems and it is assumed that the input to these cascaded stages is obtained from a suitable voltage amplifier which provides up to 1 milliwatt of driving power, at a 1-volt level, with little distortion over the desired frequency range.

The basic design objectives for high-quality audio amplifiers are discussed in terms of power output, frequency response, distortion levels, and other major performance requirements. The characteristics and capabilities of various output and driver stage configurations are compared to determine the optimum type of circuit for a given application while consideration is given to the conservative design of high-quality audio output and driver stages. Circuit configurations and performance data are shown for 10-watt, 25-watt, and 70-watt audio amplifiers that employ silicon power transistors.

The quality of an audio power amplifier is determined by its ability to provide high-fidelity reproduction of audio program material over the full range of audible frequencies. The amplifier is required to increase the power level of the input to a satisfactory output level with little distortion, and the sensitivity of its response to the input signals must remain essentially constant throughout the audio-frequency spectrum. Moreover, the input-impedance characteristics of the amplifier must be such that the unit does not load excessively and thus adversely affect the characteristics of the input-signal sources. The following paragraphs outline the design objectives for high-quality audio amplifiers with respect to these important performance characteristics.

Power-Output Capability

The amount of output power that an audio amplifier is required to develop is primarily a function of the dimensions of the room in which the unit is to be used, of the size and efficiency of the loudspeaker, and of the maximum distortion that can be tolerated at high sound levels. In general, the efficiencies of loudspeakers vary from a few percent for typical high-quality dynamic speakers down to only a few tenths of a percent for acoustic-suspension types. In typical residential-size rooms, the amplifier is required to deliver an average electrical power of several tenths of a watt to dynamic speakers at normal listening levels. Acoustic-suspension speakers, however, have a much lower transducer efficiency and may require several watts of electrical input power to produce the same sound intensity. Because high-intensity transient sounds (*e.g.*, those produced by percussion instruments) may exceed normal listening levels by a factor of ten or more, high-quality amplifiers should be capable of delivering 20 to 40 watts of output power without excessive distortion so that the transient sounds will be reproduced fully by the low-efficiency speaker. For less stringent requirements or for higher efficiency speakers, a 5- to 15-watt output capability is usually adequate.

The ability of an amplifier to supply a given amount of output power is dependent upon the loudspeaker impedance, the breakdown voltage and current-handling capability of the transistors, and the ability of the transistors to withstand the currents and voltages which may exist under all operating conditions. For all cases of practical interest, the amplifier output power, P_o, is given by the following equation:

$$P_o = \frac{I_p^2 R_L}{2} = \frac{V_p^2}{2R_L} = \frac{V_{c(max)}^2}{8R_L}$$

where I_p is the peak transistor current, V_p is the peak voltage developed across the load impedance, R_L, seen by the transistor, and $V_{c(max)}$ is the maximum allowable transistor collector voltage.

Fig. 1 shows the peak transistor current and peak transistor load voltage required to deliver a specified output power into the load impedance seen by the transistor. The curves are idealized in that the transistor saturation voltage is neglected. The transistor breakdown voltage rating must be at least twice the peak load voltage. The vertical lines denoting 4-ohm, 8-ohm, and 16-ohm resistances are particularly useful for transformerless designs, where the transistor works directly into the loudspeaker.

It is apparent from the equation that, for a given amount of output power, a high value of load impedance will require a larger voltage capability and lower current capability than is required with a low value of load impedance. Transformer-coupled output stages are frequently used to provide an optimum impedance match between the output transistors and the speaker load impedance. However, the advent of high-current power transistors which are highly resistant to failure because of *second breakdown* has made possible the design of reliable circuits without using transformer-coupled outputs.

(Second breakdown is not a voltage breakdown, but rather an electrically and thermally regenerative process in which current is focused in a very small area. The very high current, together with the voltages across the transistor, cause a localized heating that may melt a minute hole from the collector to the emitter of the transistor and thus cause a short circuit. This regenerative process is not initiated unless certain high voltages and currents are coincident for certain finite lengths of time.)

Irrespective of how the amplifier is coupled to the load impedance, the dissipation rating of the output-stage transistors must be adequate for the maximum levels in the particular circuit configuration used. In class-A amplifiers, maximum dissipation occurs when no input signal is applied; in class-B push-pull amplifiers, however, the dissipation is maximum at 42 percent of the maximum output. In addition to an adequate thermal-dissipation capability, the transistors must be able to withstand worst-case conditions of voltage and current to avoid failure from second breakdown. The worst-case operating levels for the transistors usually occur under conditions of high line voltage, at high volume levels, and at the extremes of the amplifier frequency range (*i.e.*, below 20 Hz and above 20 kHz). The operating conditions become more severe at the frequency extremes because the speaker impedances are then substantially different from their nominal value. Transistor load lines plotted under such conditions indicate that the load impedance is highly reactive.

From the viewpoint of power-dissipation rating, current and voltage capabilities, and freedom from second breakdown, a hometaxial-base transistor such as the 2N3055 appears to be an excellent choice for use in an audio output stage. Such transistors are constructed by a single-diffused technique that results in uniform junctions and a base structure having a homogeneous resistivity in the axial direction. As a result of the axially homogeneous (hometaxial) resistivity, the base structure is free from built-in axial electric fields, and the risks of second breakdown and other causes of electrical failure are thereby substantially reduced.

The main limitation on the use of the 2N3055 transistor in such applications is its moderate frequency capability. This transistor provides a gain-bandwidth product of 1 MHz at a collector current of 3 amperes; at high power levels, a frequency response that is flat within 1 dB at frequencies up to 25 kHz can be achieved. An increase in the upper limit of the frequencies for which a flat response can be obtained from the 2N3055 transistor requires a substantial amount of negative feedback. A significant increase in the drive power is then required to obtain the same level of output power.

Frequency Response

The audio-frequency range is generally considered to include those frequencies between 16 Hz and 22,000 Hz. At normal listening levels, the perception range of the average listener, however, is only from about 50 to 14,000 Hz. Moreover, the frequency response of even the best quality loudspeaker systems is usually limited to about 30 to 18,000 Hz and for most typical high-fidelity speaker systems the response range is only about 60 to 12,000 Hz. These limitations indicate that an amplifier designed so that its frequency response is flat within 1 dB from 20 to 20,000 Hz is more than adequate for any high-fidelity audio system. Several advantages are obtained, however, from the use of amplifiers having a wider frequency response.

When the frequency response of the amplifier is substantially greater than the audio spectrum, the amplifier will be stable over the audio range because the phase shift of signals in this range will be negligible. It is unlikely, therefore, that the amplifier will break into audio-frequency oscillations as a result of high-level signal transients or other causes. A wide frequency response (without feedback) is also advantageous from the standpoint of transistor dissipation, because the large phase shifts that may occur when the frequency response of the output stage is poor can also lead to excessive transistor power dissipation in both the output and driver stages.

At frequencies beyond the upper limit of the flat response, the voltage gain of the output stage decreases as frequency increases. The negative feedback voltage, which is taken from the output, is also reduced. As a result, the drive signal effectively becomes larger and it is possible that the driver stages will be overloaded. For this reason, the frequency response of the output stage should be flat over the entire range of expected signal frequencies. In fact, the over-all amplifier frequency response characteristic may be deliberately rolled off at the input or driver stage to prevent such occurrences. When the output-stage transistor has a very high frequency capability, controlled roll-off at the input is still desirable to reduce the possibility of oscillation at frequencies well beyond the audio range. The effects of amplifier frequency response on stability and dissipation are discussed in more detail in a subsequent section.

The over-all bandwidth of an audio-amplifier system is less than the bandwidth of any individual element (*e.g.*, input transducer, preamplifier, power amplifier, or loudspeaker) in the system. The bandwidth of the power amplifier should, therefore, be substantially greater than that of any of the other components so that the effect of the amplifier on the over-all response of the system is held to a minimum. For example, if the power amplifier upper-frequency 3-dB point is 50 kHz and it is driven by a preamplifier having an upper 3-dB point of 25 kHz, the 3-dB point of the combination will occur at 22 kHz. If the 3-dB point for each stage occurs at 25 kHz, the 3-dB point for the combination will occur at 16 kHz. Moreover, the frequency response will fall off more rapidly than for either of the two amplifiers by itself. This is illustrated in Fig. 2, which shows the over-all response for a power amplifier with 3-dB frequencies at 15 Hz and 30 kHz, and a preamplifier with 3-dB points at 20 Hz and 25 kHz. The examples assume that both stages have a simple *RC* response characteristic (*i.e.*, roll-off occurs at the slope of 6 dB per octave).

On the basis of the factors discussed in the preceding paragraphs, a frequency response which is flat (within 1 dB from 15 to 50,000 Hz is a reasonable design goal for the power amplifier of a very high quality audio system. A response that is flat within 1 dB from 30 to 20,000 Hz is satisfactory for a good quality system known to be stable.

Distortion

The principal types of distortion that must be considered in the design of high-fidelity audio power amplifiers are total harmonic distortion and intermodulation distortion. Both types result from nonlinearities in transistor characteristics and, in transformer-coupled circuits, from nonlinearities in the large-signal transformer characteristics. The amplifier does not contribute to the frequency distortion which may be introduced by other system components, such as loudspeakers, phonograph pickups, or tape pickups, provided that the frequency response of the amplifier is flat over the full audio range. Phase distortion, which may be introduced by factors such as improper placement of low- and high-frequency speakers, is not a major consideration in the design of audio power amplifiers. The ear is insensitive to the effects of the small audio-frequency phase shifts, even when the amplifier used has only moderate frequency-response characteristics.

Nonlinearities in the amplifier circuit cause harmonics of single-frequency signals to be generated. The *total harmonic distortion* is a measure of the magnitude of such harmonics in the amplifier output. If the input signal contains several frequency components, nonlinearities in the amplifier will result in mixing of the various input frequencies and of their harmonics to produce sum and difference frequencies of the mixed components. *Intermodulation distortion* is a measure of the effect of this mixing on the amplifier output.

In general, the *critical* listener can perceive a total harmonic distortion of about 0.7 percent and can tolerate about 2 percent. The levels of perception and tolerance, however, vary with the order of the harmonic, because the ear is more sensitive to the higher order harmonics. For example, it is possible that a listener may not notice 1 percent of second-harmonic distortion, but will find only 0.5 percent of fifth-harmonic distortion objectionable. Fortunately, the amplitude of the harmonic usually decreases as the order of the harmonic rises, and a total harmonic distortion of 0.5 percent is an acceptable upper limit for a class-B push-pull output stage, which suppresses even-order harmonics.

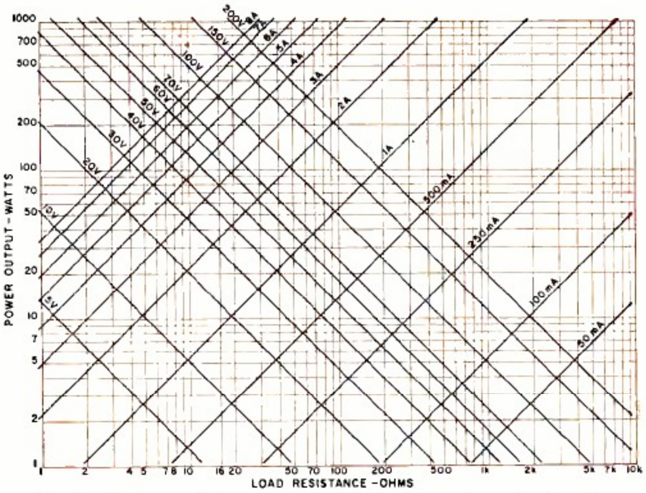


Fig. 1. Peak transistor currents and peak transistor load voltages for various output powers and load resistance values.

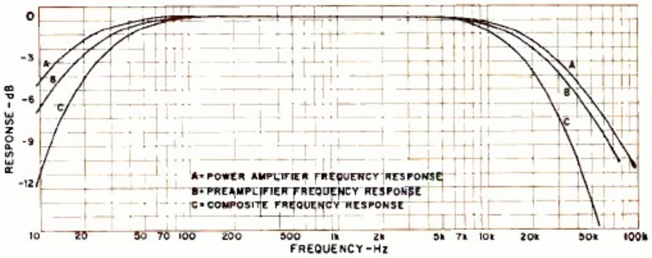
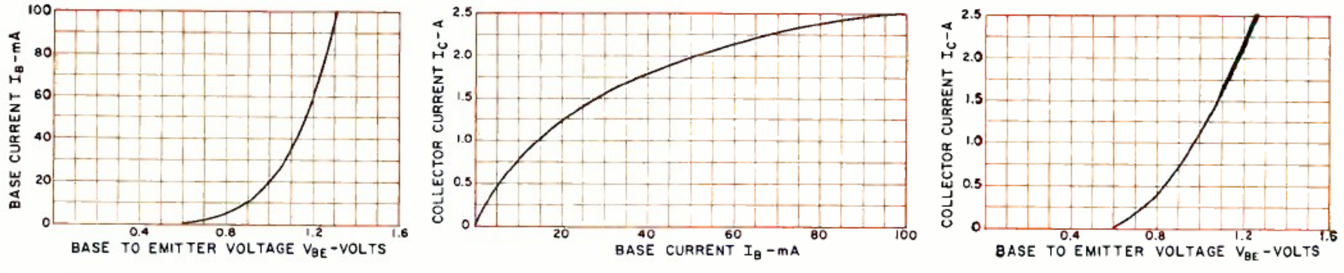


Fig. 2. The composite frequency response of the over-all audio system is poorer than the response of its poorest component.

The sensitivity of the ear to intermodulation distortion is dependent upon the amplitude and the frequency of the individual signal components. A maximum of 2 percent of intermodulation distortion is satisfactory for a two-component test signal in which one component is about ten times higher in frequency and about one-quarter the amplitude of the other component (*e.g.*, 60 and 6000 Hz mixed at an amplitude ratio of 4 to 1). The total harmonic distortion and the intermodulation distortion are both caused by amplifier nonlinearities; as a result, when one of these types of distortion is low, the other type is also low.

In common-emitter amplifiers, distortion is primarily the result of the nonlinear relationship between the base current and the base voltage (input distortion) and between the collector current and the base current (output distortion

Fig. 3. Typical transistor curves showing the nonlinearities of the input, output, and transfer characteristics.



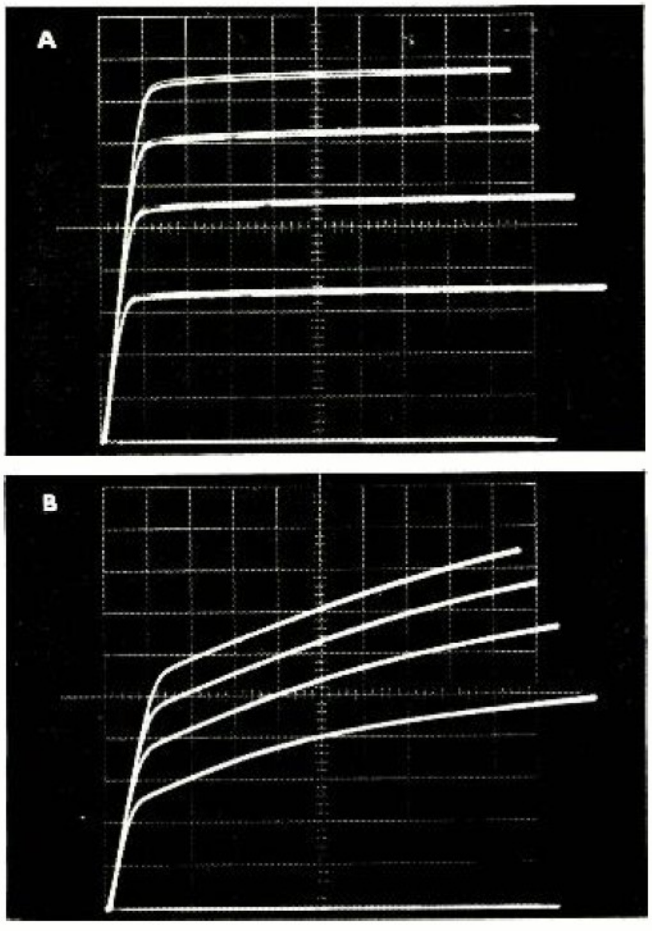
caused by nonlinear *beta*). These nonlinearities can be described in terms of the nonlinear relationship between the collector current and the base voltage (*i.e.*, nonlinear transconductance characteristics) which is useful when the driver stage has a low source impedance. Typical curves of transistor input, output, and transfer characteristics are shown in Fig. 3.

Class-B push-pull amplifiers introduce crossover distortion, which can be highly objectionable at low signal levels. This type of distortion is a result of the high degree of nonlinearity between the collector current and input voltage at low collector currents. This distortion is substantially reduced when the transistors in the push-pull stage are operated with a small quiescent current (*i.e.*, class AB).

At large signal levels, nonlinearities in transistor audio amplifiers result mainly from the dependence of the transistor *beta* on both collector current and collector voltage and from signal clipping. The reduction in *beta* at high collector currents is a well-known effect common to all transistors. The *beta* also varies nonlinearly with collector voltage in all transistors. One advantage of the hometaxial-base family of transistors is that the dependence of *beta* on collector voltage is considerably less pronounced than that in comparable diffused-base transistors. Fig. 4 compares the collector characteristics of a 2N3055 hometaxial-base transistor and a 2N3878 diffused-base transistor. The "softer" saturation characteristics of the 2N3878 transistor imply that a greater nonlinearity exists between *beta* and collector voltage in diffused-base transistors than in hometaxial-base transistors.

Signal clipping occurs whenever the transistor in an amplifier stage is driven into saturation or, for class-A stages, below cut-off. In addition, distortion may be introduced in low-cost transformer-coupled amplifiers at large signal levels because of magnetic saturation of the transformer core material.

Fig. 4. Collector characteristics of (A) 2N3055 hometaxial transistor compared with (B) 2N3878 diffused-based transistor. In both traces the vertical scale is collector current (0.5 A/div.), the horizontal scale is collector-to-emitter voltage (1 V/div.), with four different base currents (20 mA/step).



The nonlinear input and output characteristics of a transistor tend to compensate for each other over a range of collector current, as can be seen in Fig. 3. This compensatory effect results from the fact that the base current increases exponentially with base voltage and the *beta* decreases rapidly at high current levels. Optimum compensation can be achieved by the addition of a resistance in series with the base to obtain maximum linearity for the output current-input voltage transfer characteristics. The disadvantage of this technique is that a larger drive voltage is required to obtain the maximum output current and that because of variations in the characteristics of transistors, even in those from the same production run, it is difficult to determine the optimum conditions for all transistors of any given type.

Negative feedback is the most common method used to reduce distortion in audio amplifiers. In general, the feedback is taken from the output and returned to the input stage. In addition, local feedback may be employed in each stage. The amount of over-all feedback required is determined by the performance and stability objective for the amplifier. With negative feedback, the distortion can be reduced, under some conditions, to 0.1 percent or lower. An additional advantage of the negative feedback is that the output impedance of the amplifier can be reduced to a fraction of an ohm to obtain a high loudspeaker damping factor (*i.e.*, ratio of loudspeaker impedance to amplifier source impedance).

A disadvantage of the feedback is that a larger input signal is required to obtain the rated output. Moreover, when local feedback is used in a stage, steps must be taken to insure that the reduction in distortion in this stage is not offset by increased distortion in the stage that drives it because of the larger signal that the driving stage is required to supply. The feedback also either increases or decreases the input impedance of the amplifier, depending upon whether it is returned to the emitter or to the base of the common-emitter input stage that is used.

Input Impedance

In general, the input impedance of an audio power amplifier must be high enough so that the characteristics of the preamplifier in the system are not adversely affected. If the input impedance of the power amplifier is too low, the output coupling capacitor of the preamplifier (for resistance-capacitance-coupled stages) may have to be excessively large so that the low-frequency response of the system will not be degraded. Moreover, for the same input voltage, the input current of the power amplifier will be larger because of the lower input impedance. Additional distortion may then be introduced into the preamplifier because of excessive loading by the power amplifier. For good-quality audio power amplifiers, typical values of input impedance range from approximately 5000 ohms to 100,000 ohms depending upon the capability of the signal source that is used.

Voltage Sensitivity

The characteristics of the preamplifier also dictate the voltage sensitivity required of the audio power amplifier. Although a few high-quality preamplifiers have been designed which provide signal outputs of several volts, output levels in the range of 0.5 to 1.0 volt are more typical. Audio power amplifiers having a sensitivity such that the rated power output is obtained for an input of 0.5 volt and having an input impedance of 10,000 ohms are, therefore, compatible with many

(*Continued on page 67*)

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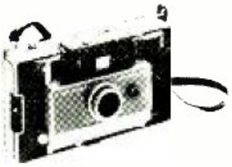

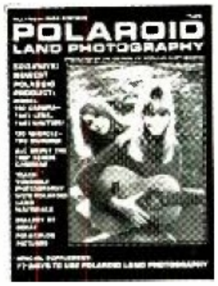

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Designing Transistor Amps

(Continued from page 26)

preamplifier designs in common use.

For increased sensitivity, the power-amplifier unit will frequently contain one or two stages of voltage amplification preceding the output stage. The power amplifier may also be driven directly from a ceramic phonograph pickup (as is done in some inexpensive table-top audio systems); up to two volts of input drive may then be available at the input of the amplifier. For operation in this way, however, the input impedance of the power amplifier should be at least 100,000 ohms and is usually 1 megohm or greater.

Hum and Noise Level

The hum level is substantially lower in transistor amplifiers than in tube amplifiers, because there is no heater supply to act as a source of the hum. Moreover, it is relatively easy to filter the ripple from transistor voltage-amplifier stages, because the current levels are low. Some precautions are necessary however to isolate the collector-supply ripple from the emitter-base terminals of each stage. Techniques used for this purpose are to be discussed in a later section.

The statistical nature of current flow through transistors, the presence of transistor and circuit resistance, and transistor surface phenomena, all result in noise sources in the amplifier. This noise is most serious in high-gain preamplifiers, where signal voltage may be only a few millivolts. The problem is much less severe for power amplifiers, because of the larger available signal voltage. It is necessary, however, to assure that the power amplifier does not contribute to the over-all system noise level. A noise level of 50 dB below 1 watt is sufficient, and levels of 60 to 70 dB below 1 watt are not difficult to achieve in a well-designed power amplifier. For example, the equivalent noise voltage for an amplifier having an input resistance of 10,000 ohms and a bandwidth of 20 kHz is less than 2 microvolts. This value is negligible compared to the input voltage required for normal, or even quiet, listening levels (*i.e.*, about 50 μV).

Next month's installment will cover various circuit configurations that are used while the concluding article will give some general conditions for conservative design and will include some practical circuits and show their performance. These particular circuits are for a 10-watt, for a 25-watt, and for a 70-watt audio power amplifier. These output powers are for a single channel of amplification.

(Continued Next Month)

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PART 2. CIRCUIT CONFIGURATIONS

Designing Silicon-Transistor Hi-Fi Amplifiers

A number of basic circuit configurations for readily available silicon power transistors, along with their comparative operating characteristics, are covered.

By R.D. GOLD and J.C. SONDERMEYER
RCA Electronic Components & Devices

LAST month we covered the performance objectives in designing hi-fi power amplifiers using silicon transistors. In this installment we will go into the circuit configurations that are used in such amplifier designs.

The type of circuit configuration selected will be dictated by the requirements of the given application. The output power to be supplied, the required sensitivity and frequency response, and the maximum distortion limits, together with the capabilities and limitations of available devices, are the main critera used to determine the circuit that will provide the desired performance most efficently and economically.

Class-A Transformer-Coupled Amplifiers

Low-cost, low-power audio systems, in which high operating efficiency is not an important consideration, usually employ a single-ended, class-A, transformer-coupled output stage, such as that shown in Fig. 1. One or more driver stages, negative feedback from the output stage to the input stage, and other features may also be necessary to obtain the desired circuit sensitivity, input impedance, and distortion characteristics.

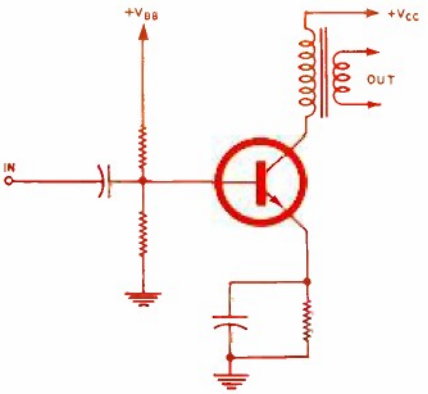


Fig. 1. Class-A transformer-coupled stage.

Fig. 2. Three-stage, transformer-coupled, class-A amplifier.

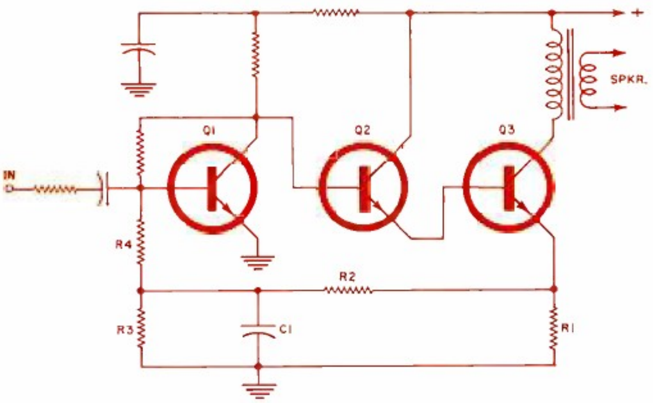


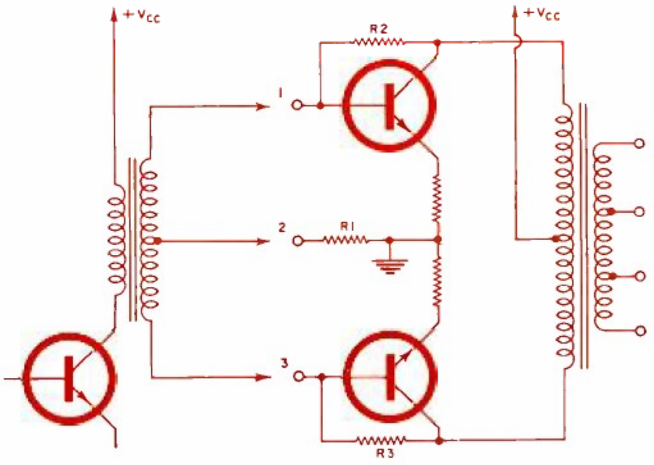
Fig. 2 shows a three-stage, class-A, transformer-coupled audio amplifier which uses d.c. feedback (coupled by $R1$, $R2$, $R3$, $R4$, and $C1$) from the emitter of the output transistor to the base of the input transistor to obtain a stable operating point. An output capability of 5 watts, with 3 percent of total harmonic distortion, is typical for this type of circuit. This output level, in general, is the upper limit for class-A amplifiers because the power dissipated by the output transistor in such circuits is more than twice the output power. For this reason, it is economically impractical to use a class-A audio amplifier to develop higher levels of output.

Class-AB Push-Pull Amplifiers

At output-power levels above 5 watts, the operating efficiency of the circuit becomes an important factor in the design of audio power amplifiers. The circuit designer may then consider a class-AB push-pull amplifier for use as the audio output stage. In such amplifiers, the operating efficiency is high (up to 78 percent), the power dissipated by the transistors under quiescent conditions is low, and, with the transistor operated class-AB (rather than class-B), crossover distortion is essentially eliminated.

Fig. 3 shows a class-AB, push-pull transformer-coupled audio output stage. Resistors $R1$, $R2$, and $R3$ form a voltage divider to provide the small amount of transistor forward bias required for class-AB operation. The transformer type of output coupling used in the circuit is advantageous in that a suitable output transformer can be selected to match the audio system to any desired load impedance, which assures a maximum transfer of the audio output power to the load circuit. This feature is especially important in sound distribution systems which use high-impedance transmission lines to reduce losses. Transformer output coupling has a major disadvantage in that it tends to limit the amplifier frequency response, particularly at the low-

Fig. 3. Class-AB, push-pull, transformer-coupled output stage.



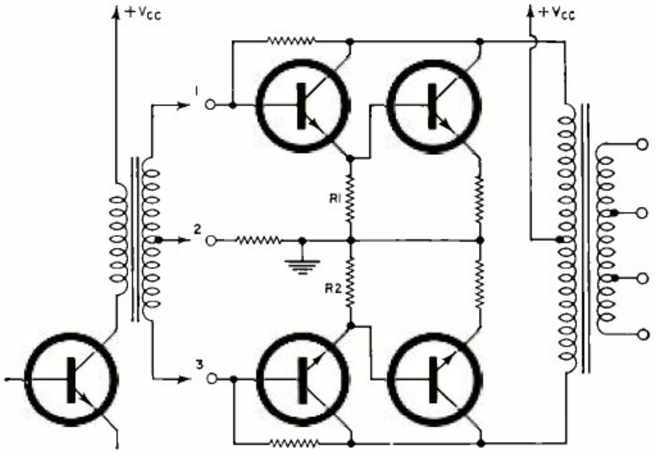


Fig. 4. Same as Fig. 3 except that Darlington pairs are employed to reduce drive requirements of output transistors.

frequency end. Variations in transformer impedance with frequency may produce significant phase shifts in the signal at both frequency extremes of the amplifier response. Such phase shifts are potential causes of amplifier instability if they occur within a feedback loop. Because of the high cost of audio-output transformers having a broad, flat frequency response, high-quality wideband transformer-coupled amplifiers cannot usually be realized economically.

Push-pull output stages, which use identical output transistors, require some form of phase inversion in the driver stage. In the circuit shown in Fig 3, a center-tapped driver transformer is used for this purpose. The requirements of this transformer depend upon the power levels involved, the bandwidth required, and the distortion that can be tolerated. This transformer also introduces phase-shift problems which tend to cause instabilities in the circuit when high levels of feedback are employed. These problems are substantially reduced when the output stage is designed to operate at low drive levels.

Fig. 4 shows a Darlington-pair output circuit used to reduce the drive requirements. Resistors *R*1 and *R*2 provide a leakage path for the driver and also permit the output transistors to turn off more rapidly. Impedance levels between the class-A driver and the output stage can be easily matched by the use of an appropriate transformer turns ratio, regardless of whether a Darlington-type of circuit is used. However, care must be taken to insure that hum pickup by the transformer is not excessive, especially when the Darlington circuit is used, because of its high gain.

An alternative method of phase inversion is to use a transistor in a phase-splitter circuit, as shown in Fig. 5A. Unlike the center-tapped transformer method, impedance matching may be a problem, because the collector of the driver, which has a relatively high impedance, operates into the low impedance of the driver stage. One solution is to reduce the output impedance of the driver stage by using smaller resistors. The resultant increase in collector current, however, also increases the dissipation. Moreover, very large coupling capacitors would be necessary to obtain good low-frequency performance. An alternate solution is to use a Darlington pair to increase the input impedance of the output stage.

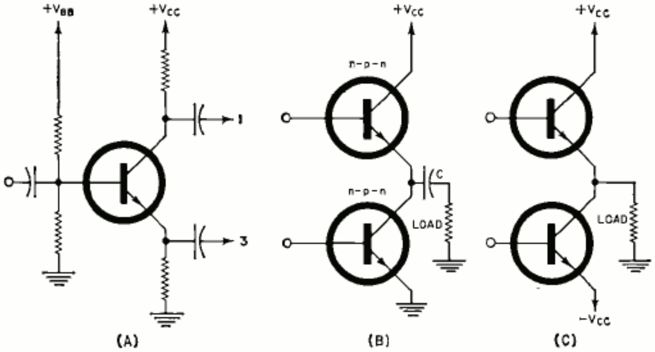


Fig. 5. (A) Transistor phase-splitter circuit. (B) Series-output circuit operated from single d.c. power supply. (C) Series-output circuit operated from opposite-polarity supply.

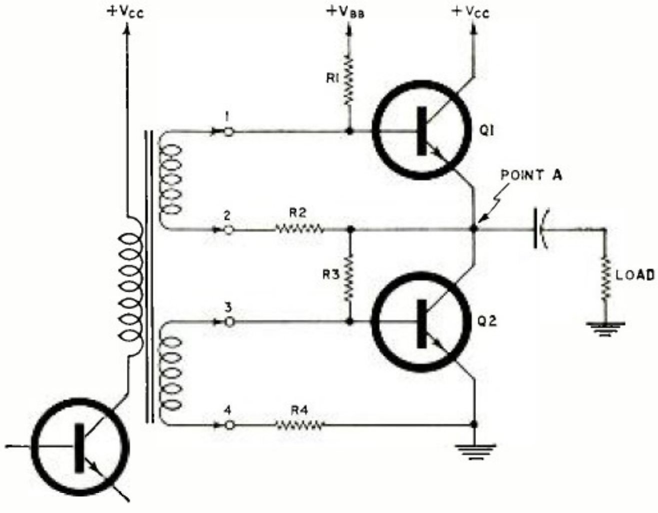


Fig. 6. Push-pull series-output amplifier with split-secondary transformer used to provide the drive-signal phase inversion.

Class-AB Series-Output Amplifiers

For applications in which low distortion and wide frequency response are major requirements, a transformerless approach is usually employed in the design of audio power amplifiers. With this approach, the most common type of circuit configuration used is the class-AB series-output amplifier in which two transistors are connected in series with the supply voltage and the load is connected to a common circuit point between these transistors. The series-output circuit may be operated from a single d.c. supply, in which case, the output is coupled to the load through a capacitor, as shown in Fig. 5B. Alternatively, the circuit may be operated from symmetrical dual (positive and negative) supplies, as shown in Fig. 5C. An advantage of the dual-supply arrangement is that there is no output coupling capacitor to limit the low-frequency response. More important, the low-frequency phase shift usually associated with such capacitors is eliminated. The requirement for two separate supplies does not necessarily increase the over-all cost of the system depending upon the choice of time constants of the power-supply filters. But additional problems may arise.

Push-Pull Driven Circuits: The class-AB-operated *n-p-n* transistors used in the series-output circuits of Figs. 5B and 5C require some form of phase inversion of the drive signal for push-pull operation. A common approach is to use a driver transformer having split secondary windings, as shown in Fig. 6. The split (separate) secondary windings are required because of the mode in which each of the series output transistors operate. If ground were used as the drive reference for both secondary windings, transistor *Q*1 would operate as an emitter-follower and would provide somewhat less than unity gain. Transistor *Q*2, however, is connected in a common-emitter configuration which can provide substantial voltage gain. In order to obtain equal output-voltage swings in both directions, the drive input to transistor *Q*1 is applied directly across the base and emitter terminals. Transistor *Q*1 is then effectively operated in a common-emitter configuration (although there is no phase reversal from input to output) and has a voltage gain that is equivalent to that of transistor *Q*2.

The disadvantages of a driver transformer pointed out previously also apply to the circuit shown in Fig. 6. In addition, coupling through inter-winding capacitances can adversely affect the performance ~~of~~ the circuit. Such coupling is particularly serious because at both ends of the upper secondary (*i.e.*, terminals 1 and 2) the a.c. voltage with respect to ground is approximately equal to the output voltage. During signal conditions when output transistor Q1 is turned on, this coupling will provide an unwanted drive to Q2. The forward transistor bias required to maintain class-AB operation of the circuit is provided by the resistive voltage divider *R*1, *R*2, *R*3, and *R*4. These resistors also assure that the output point between the two transistors (point A) is maintained at one-half the d.c. supply voltage, V_{cc}.

As in the case of the transformer-coupled output, phase inversion can be accomplished using an additional transistor. Fig. 7 shows such a circuit, in which a Darlington output stage is used to minimize loading on the phase inverter. Notice that capacitor C provides a drive reference back to the emitter of the upper output transistor. In effect, this duplicates the drive conditions of the split winding transformer approach. A disadvantage of this circuit is the high quiescent dissipation of the phase inverter (Q1) which is necessary to obtain adequate drive at full power output. An unbypassed emitter resistor (*R*) is necessary because a signal is derived from this point to drive the lower output transistor. When transistor Q1 is driven into saturation, the minimum collector-to-ground voltage that can be obtained is limited primarily by the peak emitter voltage under these conditions. In order to obtain the necessary voltage swing at this collector (which is also approximately equal to the output voltage swing), the quiescent collector-to-emitter voltage must be higher than that required in a stage that uses a bypassed emitter resistor.

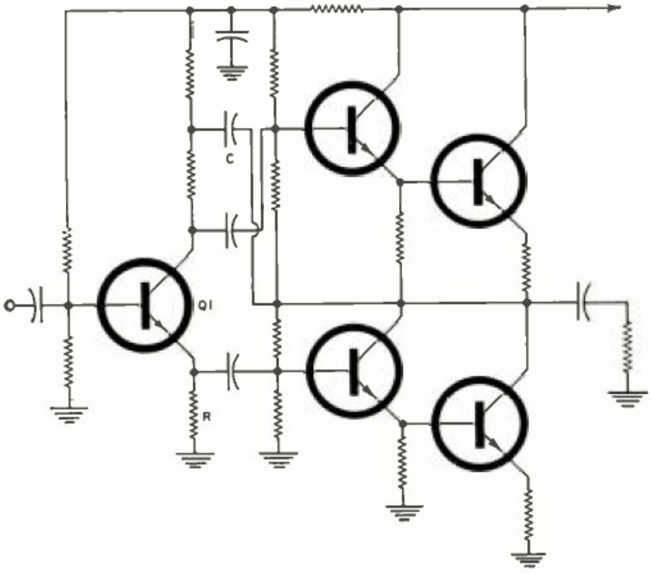


Fig. 7. Push-pull series-output amplifier in which driver and output transistors are connected as Darlington pairs and drive-signal phase inversion is provided by phase-splitter stage Q1.

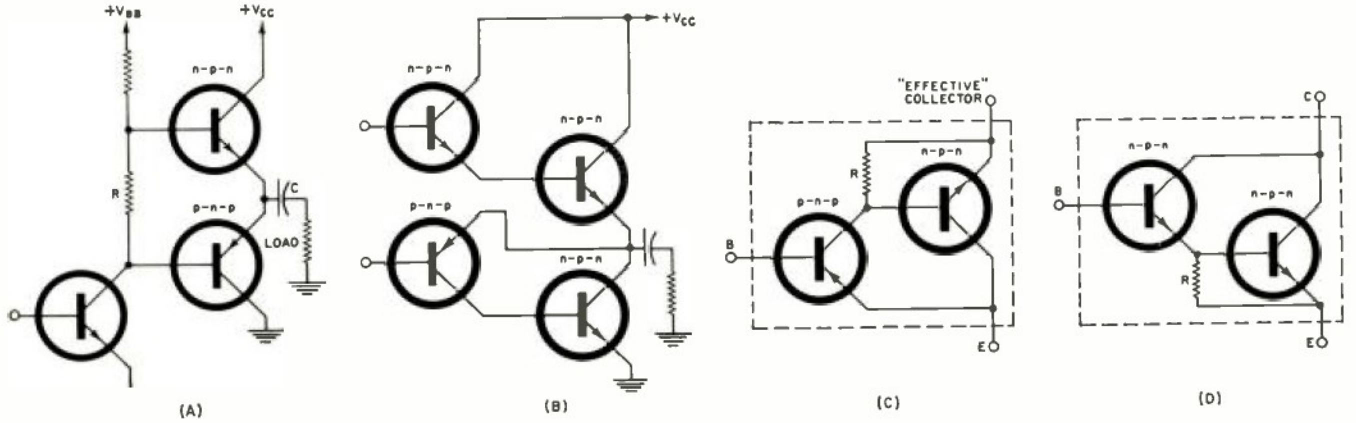
Complementary Amplifiers: With a complementary pair (*n-p-n* and *p-n-p*) of output transistors, it is possible to design a series-output type of audio power amplifier which does not require a push-pull drive. Because phase inversion is unnecessary with this type of configuration, the drive circuit for the amplifier is considerably simplified. Fig. 8A shows a basic complementary type of series-output circuit together with the simple drive arrangement that may be used. The voltage drop across resistor *R* provides the small amount of forward bias required for class-AB operation of the complementary pair of output transistors.

Rather than the use of complementary high-power output transistors, a more common approach is the quasi-complementary circuit shown in Fig. 8B. In this circuit, the driver stage employs a complementary pair of *n-p-n* and *p-n-p* transistors. This technique has become very popular with the availability of both *n-p-n* and *p-n-p* silicon driver transistors having similar electrical characteristics.

In the quasi-complementary amplifier, the driver transistors provide the necessary phase inversion. A simple but descriptive way to analyze the operation of a quasi-complementary amplifier is to consider the result of connecting a *p-n-p* transistor to a high-power *n-p-n* output transistor as shown in Fig. 8C. The collector current of the *p-n-p* transistor becomes the base current of the *n-p-n* transistor. The *n-p-n* transistor, which is operated effectively as an emitter-follower, provides additional current gain without inversion. If the emitter of the *n-p-n* transistor is considered as the "effective" collector of the composite circuit, it becomes apparent that the circuit is equivalent to a high-gain, high-power *p-n-p* transistor. The output characteristics of the *p-n-p* circuit of Fig. 8C and of a high-gain, high-power *n-p-n* circuit formed by the connection of the same type of *n-p-n* output transistor and an *n-p-n* driver transistor in a Darlington configuration, as shown in Fig. 8D, are compared in Fig. 9.

The saturation characteristics of the over-all circuit in both cases is the combination of the base-to-emitter voltage, V_{be}, of the output transistor plus the collector saturation voltage of the driver transistor. Moreover, in both cases, the current gain is the product of the individual *betas* of the transistors used. A quasi-complementary amplifier is, therefore, effectively the same as a simple complementary output circuit such as that shown in Fig. 8A, and is formed by the use of high-gain, high-power *n-p-n* and *p-n-p* equivalent transistors. In both cases, resistor *R* between the emitter and base of the output tran- (*Continued on page* 94)

Fig. 8. (A) Basic complementary type of series-output amplifier. (B) Quasi-complementary amplifier. (C) The "p-n-p" driver and "n-p-n" output transistor form the equivalent of a "p-n-p" power transistor. (D) An "n-p-n" driver and "n-p-n" transistor pair.



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CIRCLE NO. 68 ON READER SERVICE CARD

Silicon-Transistor Amplifiers

(*Continued from page* 36)

sistor reduces its storage time and also increases the effective breakdown voltage from its $V_{CEO(sus)}$ value to its $V_{CER(sus)}$ value.

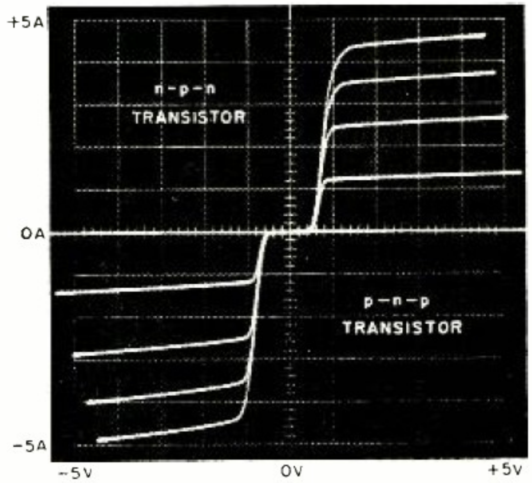
A typical quasi-complementary amplifier is shown in Fig. 10. Capacitor C performs two functions essential to the successful operation of the circuit. First, it acts as a bypass to decouple any power-supply ripple from the driver and predriver stages. Second, it is connected as a "bootstrap" capacitor to provide the drive necessary to pull the upper Darlington pair of transistors into saturation. This latter function results from the fact that the stored voltage of the capacitor, with reference to the output point "A", provides a higher voltage than the normal collector supply voltage to drive transistor $Q2$. This higher voltage is necessary during the signal conditions that exist when the upper transistors are being turned on, because the emitter voltage of transistor $Q2$ then approaches the normal supply voltage and the base voltage must be increased above this level in order to drive the transistor into saturation.

Resistor $R1$ provides the necessary d.c. feedback to maintain point "A" at approximately one-half the nominal supply voltage. Over-all a.c. feedback is coupled from output to input by resistor $R2$ to reduce distortion and to improve low-frequency performance.

As previously indicated, series-output circuits can be employed with separate positive and negative supplies, and no series output capacitor is then required. The elimination of this capacitor may result in an economic advantage, even though an additional power supply is used, because of the size of the series output capacitor necessary in the single-supply case to obtain good low-frequency performance (*e.g.*, a 2000-μF capacitor is required to provide a 3-dB point at 20 Hz for a 4-ohm load impedance). Split supplies, however, do pose certain problems which do not exist in the single-supply case. The output of the amplifier must be maintained at zero potential under quiescent conditions for all environmental conditions and devices parameter variations. Also the input ground reference can no longer be at the same point indicated in Fig. 10, because in a dual-supply system, this point is now at the negative supply potential.

If the ground-point reference for the input signal were to be the common point between the split supplies, any ripple present on the negative supply would effectively drive the amplifier through transistor $Q1$, which would operate as a common-base amplifier with its base grounded through the effective impedance of the input signal source. To avoid this

Fig. 9. Composite output V-I characteristics of Fig. 8B circuit.



condition, the amplifier must include an additional *p-n-p* transistor as shown in Fig. 11. This transistor ($Q6$) effectively reduces the drive effects of the negative supply ripple because the high collector impedance (1 megohm or higher) it presents to the base of transistor $Q1$ effectively isolates the input source impedance from transistor $Q1$. In practice, transistor $Q1$ may be replaced by a Darlington pair in order to reduce the loading effects on the *p-n-p* predriver.

Negative d.c. feedback is applied from the output to the input stage by $R1$, $R2$, and $C1$, to maintain the output at essentially zero potential. (Actually, the output will be maintained at approximately the forward-biased base-emitter voltage of transistor $Q6$, which may be objectionable in some cases, but this can be eliminated as will be shown subsequently.) Capacitor $C1$ effectively bypasses this feedback at all signal frequencies. Resistor $R3$ provides a.c. feedback to reduce distortion in the amplifier.

In next month's concluding installment, we will present some practical amplifier circuits and show their level of performance.

(Concluded Next Month)

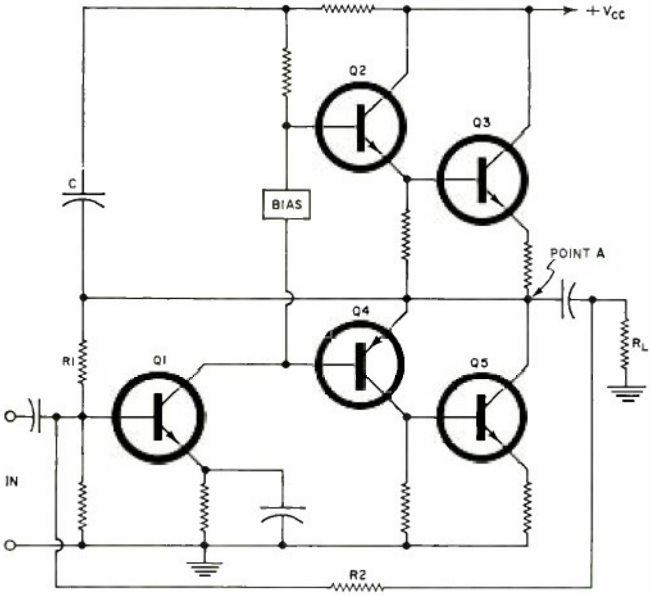
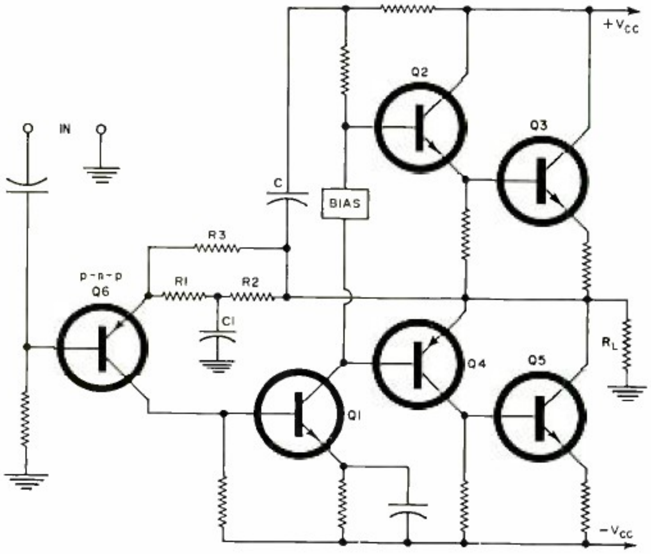


Fig. 10. A typical quasi-complementary audio power amplifier which is here shown operating from a single d.c. power supply.

Fig. 11. Quasi-complementary amplifier operating from positive and negative supplies. The "p-n-p" transistor in input stage is required to prevent ripple component from driving amplifier.



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PART 3. PERFORMANCE OF PRACTICAL CIRCUITS

Designing Silicon-Transistor Hi-Fi Amplifiers

General considerations for conservative design using readily available silicon power transistors. Practical circuits of 10-, 25-, and 70-watt amplifiers and their performance are given.

By R.D. GOLD and J.C. SONDERMEYER
RCA Electronic Components & Devices

IN addition to the consideration that must be given to the achievement of performance objectives and the selection of the optimum circuit configuration (discussed in the previous two parts of this series), the circuit designer must also take steps to insure reliable operation of the audio amplifier under varying conditions of signal level, frequency, ambient temperature, load impedance, line voltage, and other factors which may subject the transistors to either transient or steady-state high stress levels. Some of these steps are relatively straightforward. For example, it is necessary to insure that the power dissipation ratings are not exceeded at high line voltage and under worst-case signal conditions. For class-A amplifiers, the maximum power dissipation occurs at zero signal. For an ideal class-B push-pull stage, maximum power dissipation occurs when the drive signal is 64 percent of that required for maximum output power. The corresponding output power for this condition occurs at 42 percent of the maximum output power, and the dissipation in each transistor is 20 percent of the maximum power output. Also, for class-AB transformer-coupled amplifiers, the appropriate transistor breakdown-voltage rating must be greater than twice the d.c. collector voltage that is employed.

Thermal Stability Requirements

One serious problem facing the design engineer, not only in the quasi-complementary circuit but in all the circuits thus far discussed, is the ability to design a circuit which is thermally stable at all temperatures to which the amplifier might be exposed. Ideally, the quiescent current of an output stage should remain constant at all temperatures of interest. At low current levels however, the base-to-emitter voltage (V_{BE}) of a transistor decreases with increases in the junction temperature for a given collector current (I_C). If V_{BE} is held constant, then I_C will increase as the temperature rises. This behavior may lead to thermal runaway.

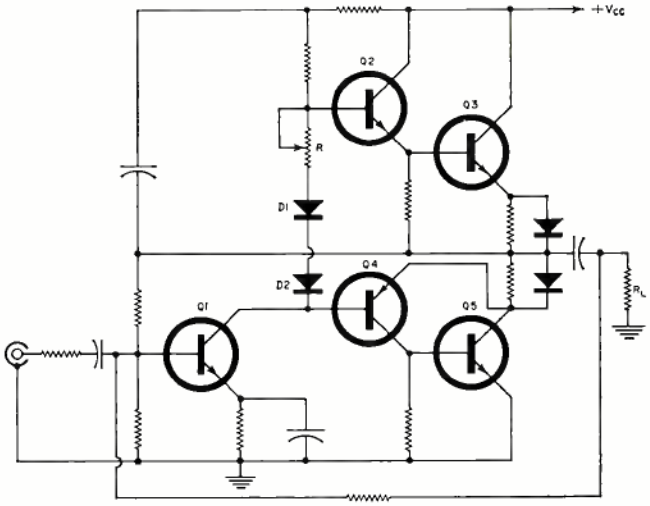
The effect of increasing temperature on collector current can be reduced by the use of an emitter resistor which will provide some local d.c. feedback. At high signal levels, the over-all saturation voltage of the device will be increased because of the voltage drop across this resistor. One solution to the saturation-voltage problem is to bypass the emitter resistor with a capacitor. In high-power amplifiers, however, the emitter resistors employed usually have a value of about 1 ohm, and the size of the capacitor required to bypass the emitter adequately at all frequencies of interest makes this approach economically impractical. A more practical solution is to increase the value of the emitter resistor and shunt it with a diode. With this technique, sufficient degeneration is provided to improve circuit stability, but the maximum voltage drop across the emitter resistor is limited to the forward voltage drop of the diode.

Thermal stability can be further improved by the addition of devices such as thermistors or bias diodes, the characteristics of which are such that they will tend to reduce the base drive voltage of the output transistor as temperatures rise. When these types of devices are used, it is possible to reduce or even eliminate emitter networks completely and thereby to reduce substantially the circuit losses at high power levels. It is interesting to note that a simple emitter diode itself will provide some improvement in circuit stability. The static resistance of a diode is fairly high at low currents (about 30 ohms at 20 milliamperes for a 1N1612). A disadvantage of this technique is that the forward voltage drop of this diode decreases with increasing temperature and, therefore, reduces the stabilizing effects of the high dynamic resistance.

It should be noted that at high current levels, the base-to-emitter voltage of silicon transistors increases with a rise in the junction temperature. This characteristic is the result of the increase in the small base resistance that is produced by the rise in temperature. The increase in base resistance with temperature has two beneficial effects: First, it helps to stabilize the transistor against thermal runaway because higher temperatures now require an increase in V_{BE} to cause an increase in I_C. Second, the increased resistance causes a portion of the transfer characteristic to be linear. A lower distortion is therefore possible at high temperatures.

The quasi-complementary amplifier shown in Fig. 1 incorporates the stabilization techniques just described. A

Fig. 1. Quasi-complementary amplifier with compensating diodes.



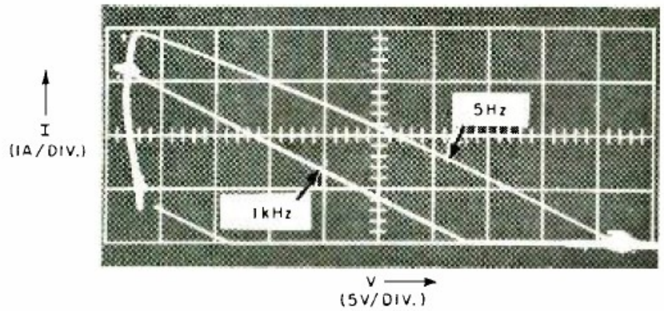


Fig. 2. Note load-line shift at very low operating frequency.

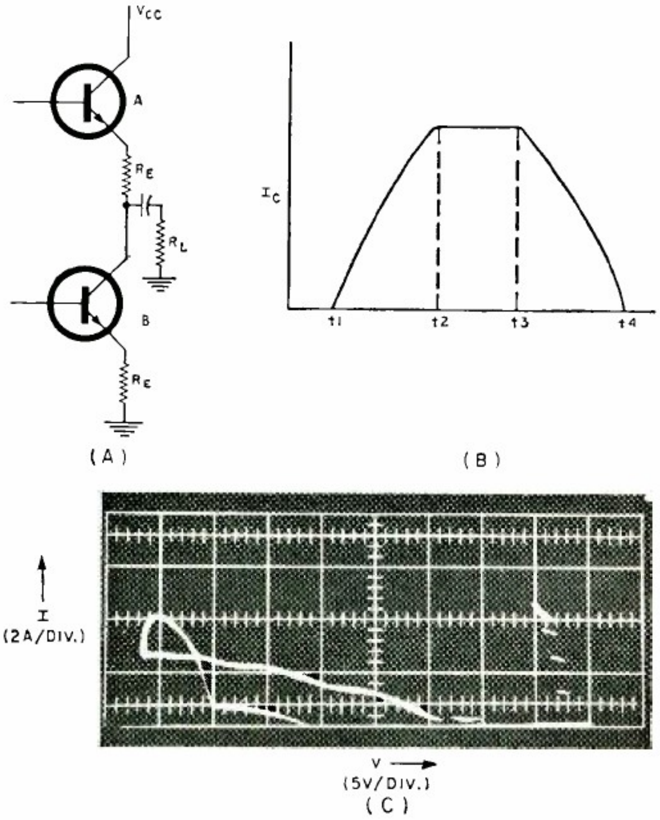


Fig. 3. (A) Output circuit which is overdriven. (B) Transistor A collector-current waveform. (C) Load line of circuit shown at operating frequency of 100 kHz and power output of 30 watts.

resistor-diode network is used in the emitter of transistor Q3, and another such network is used in the collector of transistor Q5.

Previous discussion regarding the *p-n-p* driver and *n-p-n* output combination (Q3 and Q5) revealed that the collector of the output device becomes the "effective" emitter of the high-gain, high-power *p-n-p* equivalent, and *vice versa*. Therefore, in order to provide maximum operating-point stability, the diode-resistor network should be in the "effective" emitter of the *p-n-p* equivalent. Most quasi-complementary circuits employ the stabilization resistor in the emitter of the lower output transistor and thus do not improve the operating-point stability of the over-all circuit. The resistor, however, does provide some protection against thermal runaway of the lower output transistor. Such protection may be necessary unless it is provided by other means.

The circuit shown in Fig. 1 is biased for class-AB operation by the voltage obtained from the forward drop of two diodes, *D*1 and *D*2, plus the voltage drop across potentiometer *R*, which affords a slight adjustment in the value of the quiescent current. The current necessary to provide this voltage reference is the collector current of driver transistor Q1. The diodes may be thermally connected to the heat sink of the output transistors so that thermal feedback will be provided to further improve thermal stability. Because the forward voltage of the reference diodes decreases with increasing temperature, these diodes effectively compensate for the decreasing V_{BE} of the output transistors by reducing the external bias applied. In this way, the quiescent current of the output stage can be held relatively constant over a wide range of operating temperatures.

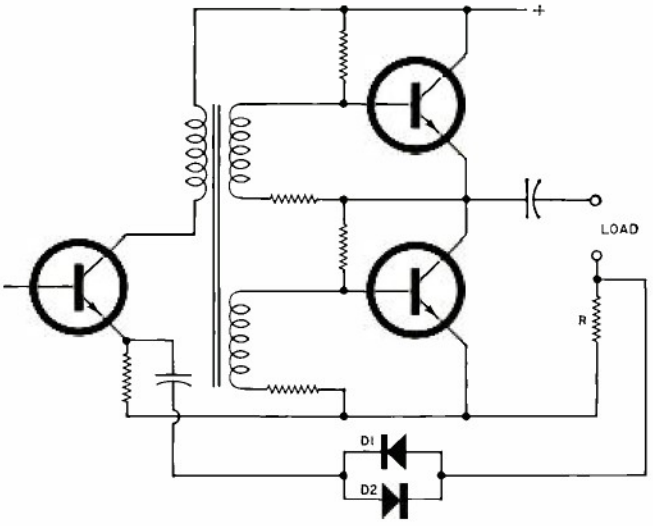
The value of the transistor operating parameters that affect thermal stability can be calculated to insure freedom from thermal runaway. In these calculations it should be realized that the temperature-dependent collector leakage-current limit specified by the transistor manufacturer actually consists of two components. One is related directly to the collector junction saturation current and is a strong function of temperature. In silicon transistors, this component is approximately doubled with each 7° C rise in junction temperature. At room temperature, however, it is on the order of only a few nanoamperes, so that a rise in case temperature of 140° C will cause the saturation current to rise only a few milliamperes.

The other component of collector leakage current is a surface leakage which is relatively independent of temperature. In fact, this leakage component may decrease as the temperature increases. The value of total leakage current (I_{CER}) specified by the transistor manufacturer is the sum of these two components. If the specified value is on the order of a few milliamperes, it will remain substantially constant with temperature. For example, in the published data for the *RCA*-40363, I_{CER} is given as 0.5 milliampere (maximum) at $T_c = 150°$ C. The transistor is, therefore, quite stable thermally with respect to any changes that might occur in the amount of leakage current.

Effects of Large Phase Shifts

The frequency-response characteristic is an important factor with respect to the ability of the amplifier to withstand unusually severe electrical stress conditions. For example, under certain conditions of input signal amplitude and frequency, the amplifier may break into high-frequency oscillations which can lead to destruction of the output transistors, the drivers, or both. This condition is particularly a problem in transformer-coupled amplifiers because the characteristics of transformers depart from the ideal at both low and high frequencies. The departure occurs at low frequencies because the transformer inductive reactance decreases and, at high frequencies because the effects of leakage inductance and of transformer wind-

Fig. 4. Push-pull power amplifier with short-circuit protection.



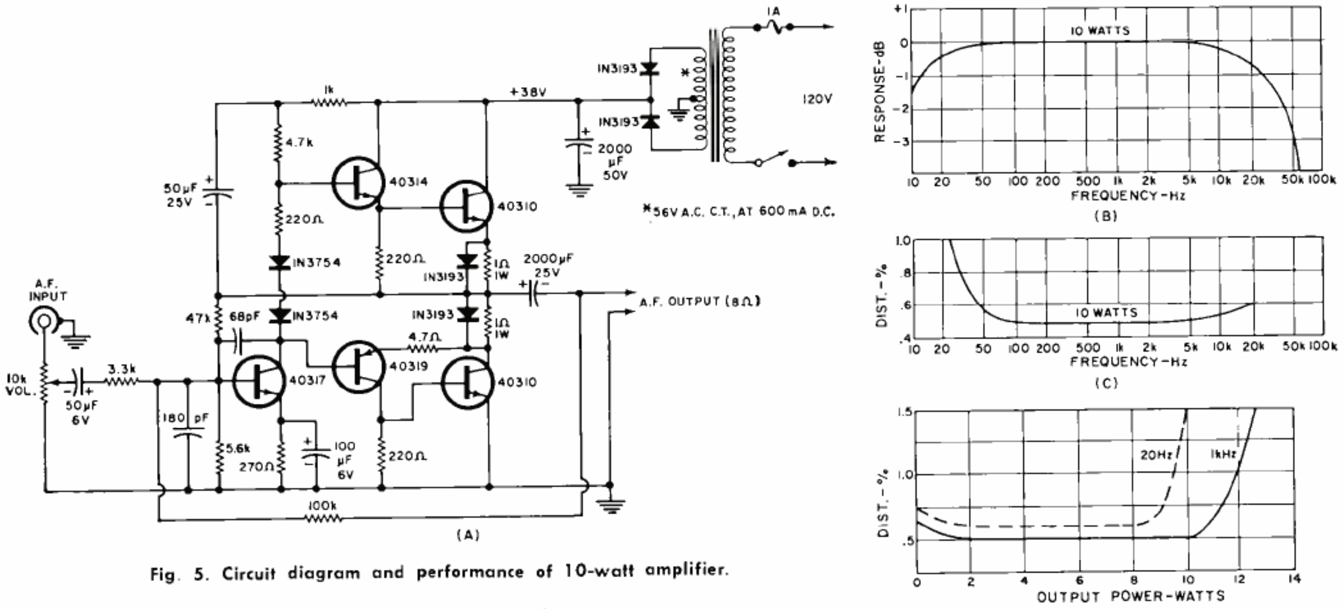


Fig. 5. Circuit diagram and performance of 10-watt amplifier.

ing capacitance become appreciable. At both frequency extremes, the effect is to introduce a phase shift between input and output voltage.

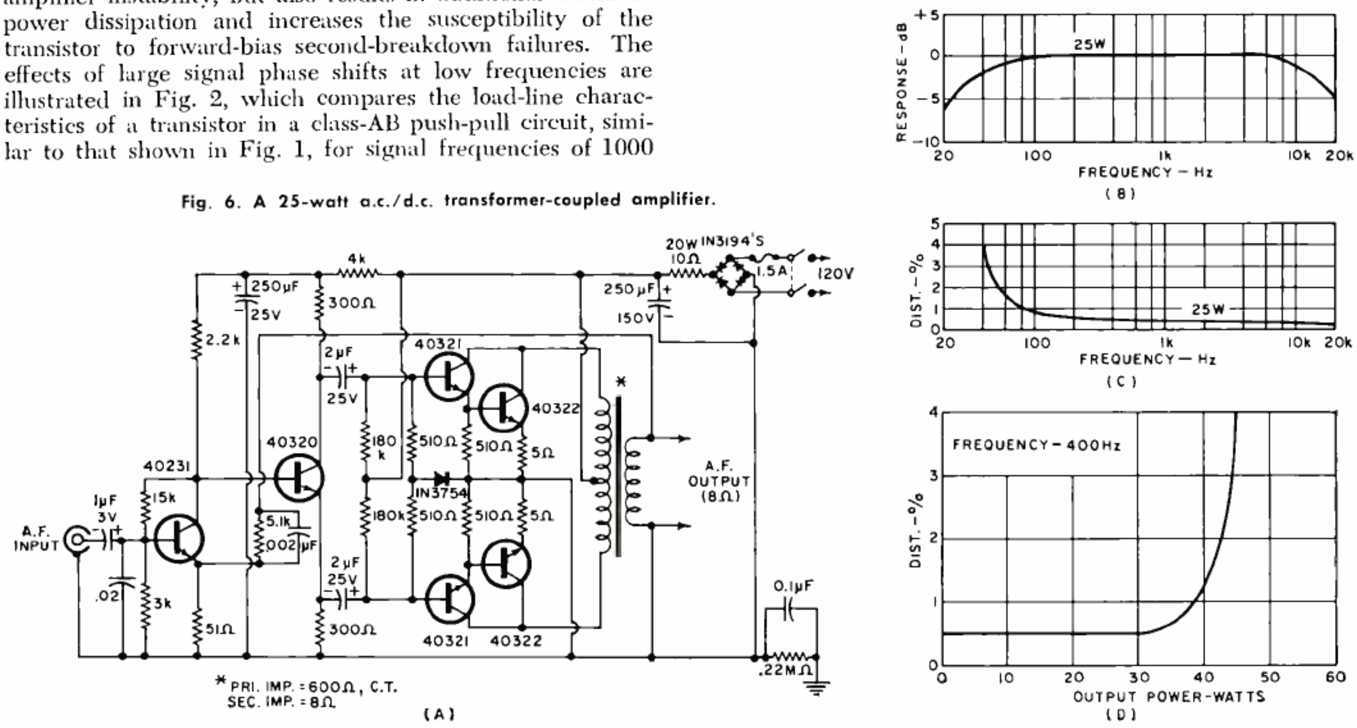
Negative feedback is used almost universally in audio amplifiers, and the voltage coupled back to the input by the feedback loop may cause the amplifier to be potentially unstable at some frequencies, if the additional phase shift is sufficient to make the feedback positive. Similar effects can occur in transformerless amplifiers because reactive elements, such as coupling and bypass capacitors, transistor junction capacitance, stray wiring capacitance, and inductance of the loudspeaker voice coil, are always present. The values of some of the reactive elements (*e.g.*, transistor junction capacitance and transformer inductance as the core nears saturation) are functions of the signal level, and coupling through wiring capacitance and unavoidable ground loops may also vary with the signal level. As a result, an amplifier which is stable under normal listening levels may break into oscillations when subjected to high-level signal transients.

A large phase shift is not only a potential source of amplifier instability, but also results in additional transistor power dissipation and increases the susceptibility of the transistor to forward-bias second-breakdown failures. The effects of large signal phase shifts at low frequencies are illustrated in Fig. 2, which compares the load-line characteristics of a transistor in a class-AB push-pull circuit, similar to that shown in Fig. 1, for signal frequencies of 1000 Hz and of 5 Hz. The phase shift is caused primarily by the output capacitor. In both cases the amplifier is driven very hard into saturation by a 5-volt input signal. The increased dissipation at 5 Hz compared to that obtained at 1000 Hz results from simultaneous high-current and high-voltage operation. The transistor is required to handle safely a current of 0.75 ampere at a collector voltage of 40 volts for an equivalent pulse duration of about 10 milliseconds; it must be free from second breakdown under these conditions of operation.

Excessive Drive Levels

Simultaneous high-current and high-voltage operation may also occur in class-B amplifiers at high frequencies when the amplifier is overdriven to the point where the output signals are clipped. For example, assume that the input signal applied to the series-output push-pull circuit shown in Fig. 3A is large enough to drive the transistors into both saturation and cut-off. During a portion of the input

Fig. 6. A 25-watt a.c./d.c. transformer-coupled amplifier.



cycle, therefore, transistor A will be driven into saturation, and transistor B will be cut off. Fig. 3B shows the collector-current waveform for transistor A under these conditions.

During the interval from $t2$ to $t3$, transistor A operates in the saturation region, and the output voltage is clipped. The effective negative feedback is then reduced because the output voltage does not follow the sinusoidal input signal. Transistor A, therefore, will be driven even further into saturation by the unattenuated input signal. When transistor B starts to conduct, transistor A cannot be turned off immediately because the excessive drive has resulted in a large storage time. As a result, transistor B is required to support essentially the full supply voltage (less only the saturation voltage of transistor A and the voltage drop across the emitter resistors, if used), as its current is increased by the drive signal. For this condition, a large input signal is required when the frequency is high enough so that the storage time is greater than one-quarter cycle.

Fig. 3C shows the type of load line obtained under such conditions. The duration of the high-current, high-voltage condition is usually short enough so that forward-bias second breakdown does not occur. For example, the load line shown is for a 2N3878 transistor operated at a frequency of 100 kHz; no second breakdown failure occurred.

Transistor A in Fig. 3A is also subject to forward-bias second breakdown if the d.c. supply voltage and a large input signal are applied simultaneously, because of the charging current through the output coupling capacitor.

If the load of a transformer-coupled amplifier is disconnected during operation, the transistor then sees an inductive load (the transformer primary inductance). When the transistor is turned off, reverse-bias second breakdown may occur. Direct- or capacitive-coupled circuits, on the other hand, are quite stable with the load removed.

If the amplifier high-frequency response is limited by the high-frequency capability of the output transistors, then the driver transistors may be unduly stressed under high-frequency, high-drive conditions. This stress is produced because the reduction in output voltage, as amplifier gain decreases, results in a smaller negative feedback voltage. The effective over-all amplifier gain is therefore increased, thereby causing the current in the driver transistors to increase. At sufficiently high frequencies, failure may then result because the drivers become overloaded.

This potential cause of failure can be avoided by the deliberate introduction of a frequency roll-off at the input, or by the use of high-frequency output transistors. The 2N3878, which has a typical gain-bandwidth product of 100 MHz, is well suited as an output for very wide band amplifiers. This transistor has been used in a low-distortion amplifier to obtain a frequency response which is down only 1 dB at 15 Hz and at 200 kHz at 20 watts output.

Short-Circuit Protection

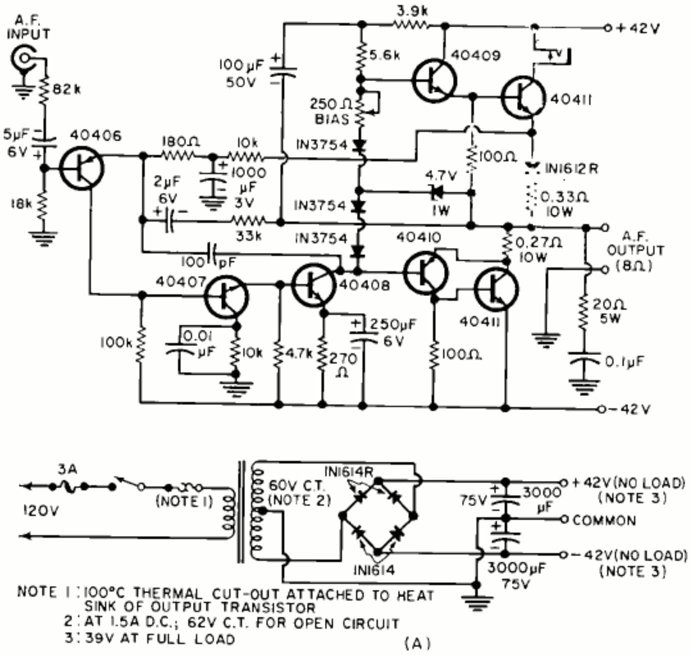
Another important consideration in the design of high-power audio amplifiers is the ability of the circuit to withstand short-circuit conditions. When the output terminals of an amplifier are shorted, the feedback becomes ineffective, and the open-loop gain is such that overdrive conditions result in disastrously high currents and excessive dissipation in both driver and output stages. Generally, before the output fuse can blow, the transistors are destroyed. Obviously, some form of short-circuit protection is necessary.

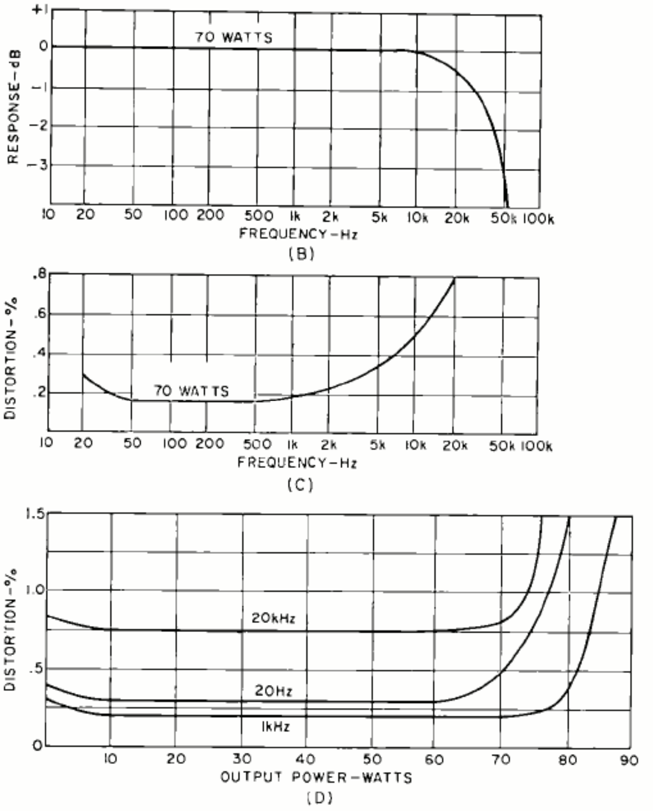
One such technique is shown in Fig. 4. A current-sampling resistor R is placed in the ground leg of the load. If any condition (including a short) exists such that higher-than-normal load current flows, diodes $D1$ and $D2$ conduct on alternate half cycles and, thereby, provide a high negative feedback which effectively reduces the drive of the amplifier; however, this feedback should not exceed the stability margin of the amplifier. Notice that this technique does not in any way effect the normal operation.

10-Watt, Class-AB Audio Amplifier

The advantages of using silicon power transistors in the driver and output stages of high-power audio amplifiers are shown by the typical performance of three practical circuits designed to operate at widely different power-output levels (10 watts, 25 watts, and 70 watts). The performance data shows that silicon transistors can be used to develop high levels of audio output power in circuits that exhibit the wide frequency response, high sensitivity, and low distortion levels required in high-quality audio systems. Moreover, because of the high-tempera- (*Continued on page* 80)

Fig. 7. A 70-watt direct-coupled audio power amplifier.





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analysis to ensure that the unit complies with the original "black-box" specifications.

Production Testing

Because an integrated unit performs a circuit or multiple-circuit function, testing can be a significant portion of the cost. However, a large part of the testing cost can be avoided by use of a.c. or dynamic specifications which permit performance deviations within the normal process variables. Under such conditions, an integrated circuit can generally be evaluated completely by means of static or d.c. testing. These static tests are fast (typical automatic-equipment static test rates are on the order of 30 to 60 tests per second) and provide a high probability that the circuit will perform properly under dynamic conditions because the components of the integrated circuit are frequently more sensitive to d.c. than a.c. evaluation.

There are obvious exceptions for which d.c. tests cannot assure compliance with an a.c. specification, particularly when capacitor coupling is used. In such cases the pellet is d.c.-probed before dicing at internal points and then may be a.c.-tested after the mounting and bonding have been completed.

The Future

The most important single development in linear integrated circuits over the next few years should be the design of standard types and the acceptance of these types by equipment manufacturers. Large volume requirements with extreme emphasis on low cost will encourage the development of special circuits with multi-function uses, particularly in consumer and some limited industrial applications. Technological developments will extend the frequency range through u.h.f., and large improvements will be forthcoming in low-noise applications at both high and low frequencies. Extension into greater power-handling capability will, to a large extent, be determined by economic considerations. Combinations of MOS and bipolar transistors offer some unique technical advantages and will become a part of the standard types offered by more manufacturers.

The most significant contribution that integrated circuits will make to equipment of the future is that more complex instruments will be generated because one or more orders of magnitude of electronic functions will be purchased for the same cost. In addition, the improvement in reliability will permit such designs. As a result, equipment manufacturers will be stimulated to design products which today are neither practical nor economically feasible. ▲

Designing Hi-Fi Amplifiers

(*Continued from page 50*)

ture capabilities of the silicon transistors, the performance of the amplifiers remains essentially the same over a wide variation in ambient temperature.

Fig. 5A shows a schematic of a practical 10-watt quasi-complementary audio amplifier. This circuit employs the same stability techniques as those used in the circuit of Fig. 1. Two 1N3754 diodes are used in the input of the driver stage to compensate for the effect of high-temperature variations of the output transistors. Two 1-ohm resistors are placed in the output stage to provide the degeneration required for circuit stability. These resistors are shunted by 1N3193 diodes to reduce losses when the amplifier is operated at full rated output power.

The use of direct-coupled stages and local d.c. feedback results in very stable quiescent operation at ambient temperatures up to 71°C. With an over-all negative feedback of 5 dB, the amplifier has a response that is flat within 1 dB from 15 to 20,000 Hz. Performance curves for the 10-watt amplifier are shown in Figs. 5B, 5C, and 5D.

25-Watt, Class-AB Audio Amplifier

Fig. 6A shows the schematic of a 25-watt a.c./d.c. transformer-coupled audio amplifier intended primarily for public-address systems and other applications for which economy and flexibility with respect to load impedance are important considerations. The high breakdown voltage of the silicon power transistors used in the output and driver stages permit the amplifier to be operated directly from a 120-volt a.c. or d.c. line. The negative-voltage terminals of the amplifier (*i.e.*, circuit ground) is isolated from chassis ground by a 0.22-megohm resistor to reduce the risks of electrical shock. The signal input should be transformer-coupled to the power amplifier to avoid shock hazard from the signal-source ground. A 0.1-μF capacitor provides a low impedance connection between circuit ground and chassis ground at r.f. frequencies to prevent high-frequency oscillations.

Each driver transistor is connected to the associated output transistor in a Darlington arrangement; the output is transformer-coupled to the speaker. Drive-signal phase inversion is provided by a transistor phase-splitter circuit. The small amount of forward bias required for class-AB operation is provided by the 180,000 and 510-ohm resistors and the 1N3754 diode. The diode also provides the temperature compensation required so that the quiescent current will remain relatively constant for wide variations in temperature. With the 10 dB of over-all feedback from the output to the emitter of the first stage, the amplifier has an input impedance of 2500 ohms. Performance curves for the 25-watt unit are shown in Figs. 6B, 6C, and 6D.

70-Watt, Class-AB Audio Amplifier

Fig. 7A shows the schematic of a high-quality 70-watt direct-coupled series-output audio amplifier in which unique techniques are used to obtain stable and reliable performance. The three 1N3754 diodes in the driver stage are thermally connected to the output transistor heatsinks so that the thermal feedback required to maintain a preset 20 milliamperes of quiescent output current is obtained at all case temperatures up to 100°C. Small-value emitter resistors are employed in the output stage because additional stability is not necessary and output losses must be held to a minimum. A 1N1612R diode is placed in the emitter of one output transistor to cancel the offset voltage of the input transistor and thereby maintain the quiescent output voltage near zero.

Short-circuit protection is provided by the 0.27- and 0.33-ohm emitter resistors and the zener diode. If any condition exists which will cause higher-than-normal current (5 amperes) to flow through these resistors, the voltage potential across the zener diode will be such that the diode conducts in the forward direction during the negative output half cycle and exceeds the diode breakdown voltage during the positive half cycle. In this way, the driver is clamped below the 5-ampere level, and no increase in output current above this value is allowed. The drivers and output transistors, therefore, are protected from high currents and excessive power dissipation that may result from a reduced load resistance or, in the worst case, a short-circuit. In addition, a 100°C thermal cut-off is attached to the output transistor heatsink which will turn off the amplifier when these abnormal conditions cause sustained higher-than-normal output dissipation.

The frequency response of the amplifier is flat within 1 dB from 5 to 25,000 Hz. The input sensitivity of the amplifier is 0.8 volt r.m.s. for full rated output. The input resistance is 100,000 ohms. The performance curves for the 70-watt amplifier are shown in Figs. 7B, and 7D.

(Editor's Note: The three amplifier circuits shown in this article are not intended as construction projects. We have no information on sources for any of the special parts required. Rather the circuits were included to illustrate the various design principles discussed in this three-part series of articles.) ▲

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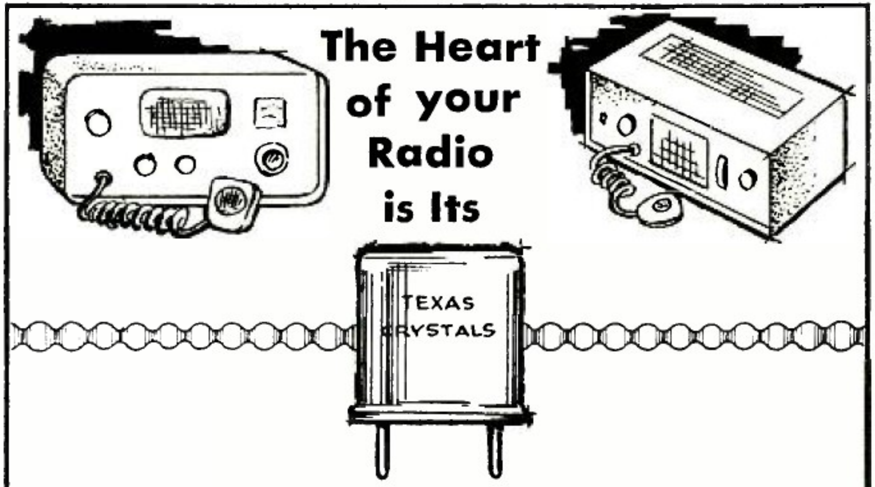

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